NASA TECHNICAL NOTE

NASA TN D-8327 C.1

NASA TN D-8327

A UNIFIED ACQUISITION SYSTEM FOR ACOUSTIC DATA

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NATIONAL AERONAUTICS AND SPACE ADMINISTRATION · WASHINGTON, D. C. · MARCH 1977

1. Report No. NASA TN D-8327	2. Government Accession No.	3. Recipient's Catalog No.
4. Title and Subtitle A UNIFIED ACQUISITION SYSTEM FOR ACOUSTIC DATA		5. Report Date March 1977
		6. Performing Organization Code
7. Author(s) Allan J. Zuckerwar and Harlan K. Holmes		8. Performing Organization Report No. L-10915
		10. Work Unit No. 505-03-11-06
9. Performing Organization Name and Address NASA Langley Research Center Hampton, VA 23665		11. Contract or Grant No.
		13. Type of Report and Period Covered Technical Note
12. Sponsoring Agency Name and Address National Aeronautics and Space Administration Washington, DC 20546		14. Sponsoring Agency Code

15. Supplementary Notes

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16. Abstract

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17. Key Words (Suggested by Author(s)) Noise measurement system Field noise measurement Sonic-boom measurement system Acoustic data acquisition system Aircraft noise measurement system	18. Distribution Statement Unclassified - Unlimited Subject Category 35

19. Security Classif. (of this report)	20. Security Classif. (of this page)	21. No. of Pages	22. Price*
Unclassified	Unclassified	28	$4.00

*For sale by the National Technical Information Service, Springfield, Virginia 22161

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SUMMARY

A multichannel, acoustic AM carrier system has been developed for a wide variety of applications, particularly for aircraft-noise and sonic-boom measurements. Each data acquisition channel consists of a condenser microphone, an acoustic signal converter, and a Zero Drive® amplifier, along with peripheral supporting equipment. A control network insures continuous optimal tuning of the converter and permits remote calibration of the condenser microphone. With a 12.70-mm (1/2-in.) condenser microphone, the converter/Zero Drive® amplifier combination has a frequency response from 0 Hz to 20 kHz (-3 dB), a dynamic range exceeding 70 dB, and a minimum noise floor of 50 dB (ref. 20 μPa) in the band 22.4 Hz to 22.4 kHz. The system requires no external impedance matching networks and is insensitive to cable length, at least up to 900 m (3000 ft). System gain varies only ±1 dB over the temperature range 4° to 54° C (40° to 130° F). Adapters are available to accommodate 23.77-mm (1-in.) and 6.35-mm (1/4-in.) microphones and to provide 30-dB attenuation. A field test to obtain the acoustical time history of a helicopter flyover proved successful.

INTRODUCTION

A unique, new instrumentation system for the acquisition of acoustic data has been developed and is described in this report. This development has resulted from past experiences in field measurement of acoustic data for, typically, fan-jet-powered aircraft, turbine-powered helicopters, and supersonic vehicles as well as in wind-tunnel investigations of turbulent boundary layers. The extremely broad frequency range and the large signal-amplitude dynamic range implicit in these applications have previously required various measurement systems, each optimized for the particular measurement. To illustrate, figures 1, 2, and 3 are representative time histories and amplitude spectra from a commercial fan-jet transport, a turbine-powered helicopter, and a sonic boom from a supersonic vehicle, respectively. The time histories show that the transport noise is relatively broadband; the helicopter noise has strong impulsive spikes; and the sonic boom has a transient signature with the typical N-wave shape. The amplitude spectrum of the transport possesses a broadband characteristic with fan and compressor tones superimposed. The helicopter signature possesses a low-repetition-rate blade frequency and its many harmonics superimposed on a broadband spectrum, with the tail rotor noise evident at the higher frequencies. The sonic-boom signature is described by infrasonic frequencies (governed by the relatively long separation between the sharp acoustic pressure transitions) and the higher harmonics characteristic of transient events.

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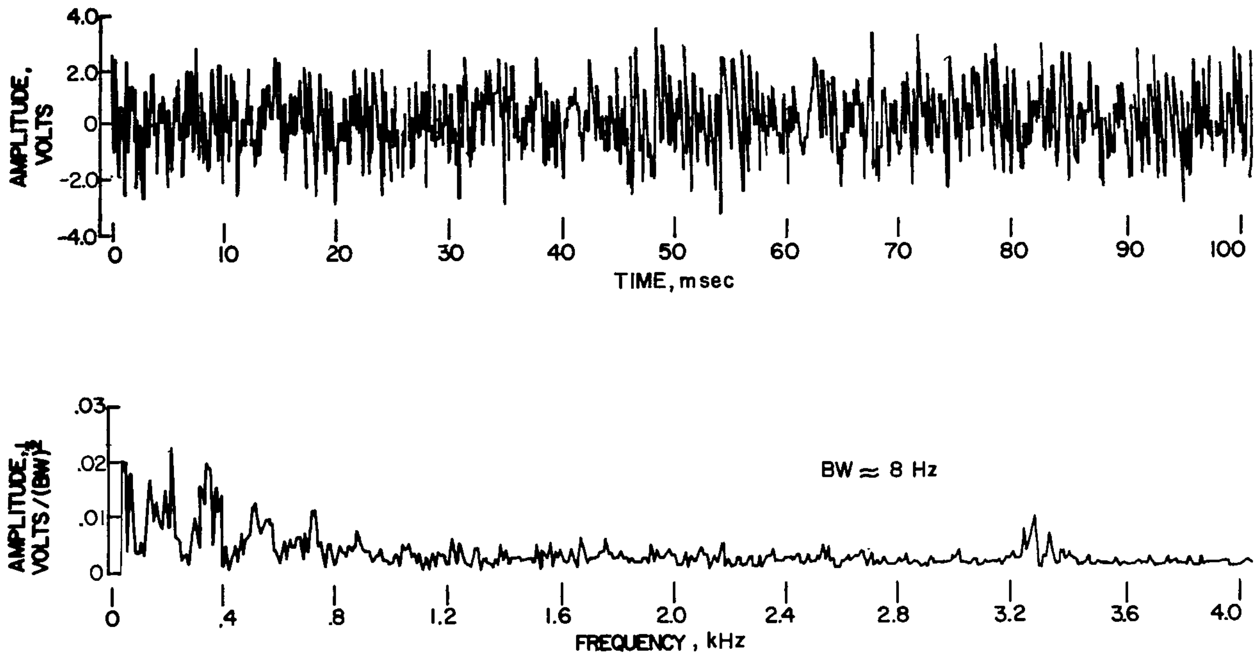

Figure 1.- Amplitude time history and narrow-band spectrum of a commercial fan-jet transport.

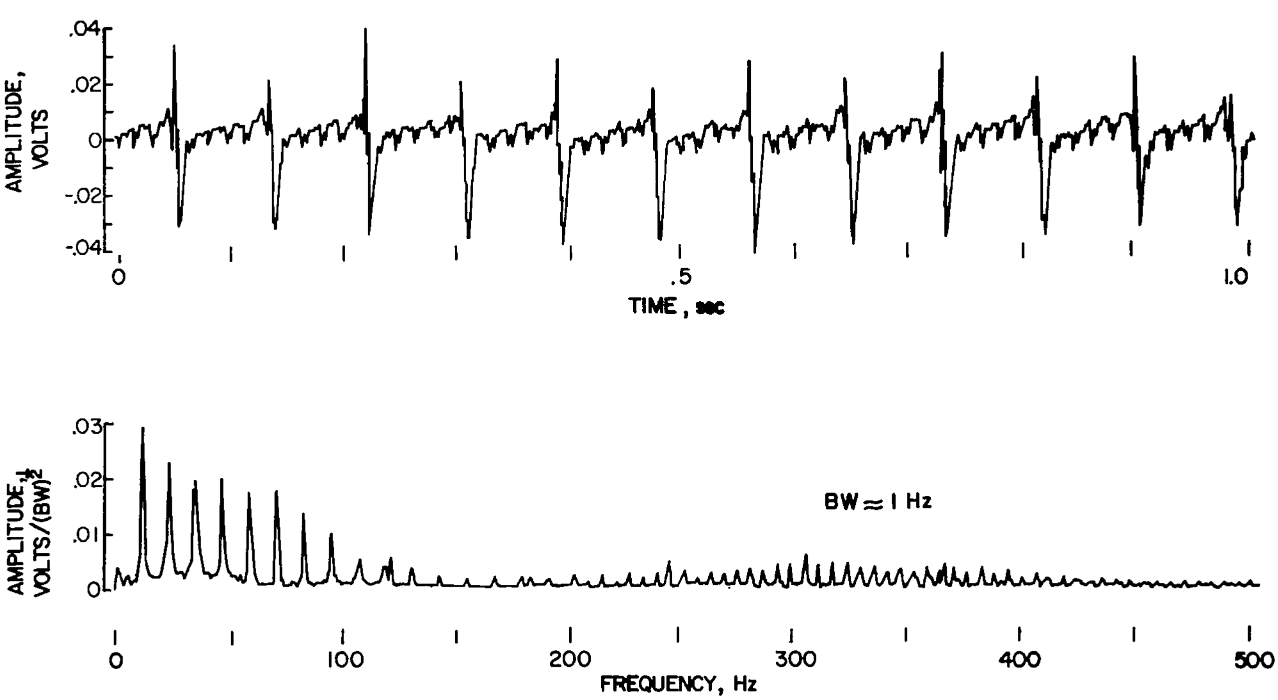

Figure 2.- Amplitude time history and narrow-band spectrum of a turbine-powered helicopter.

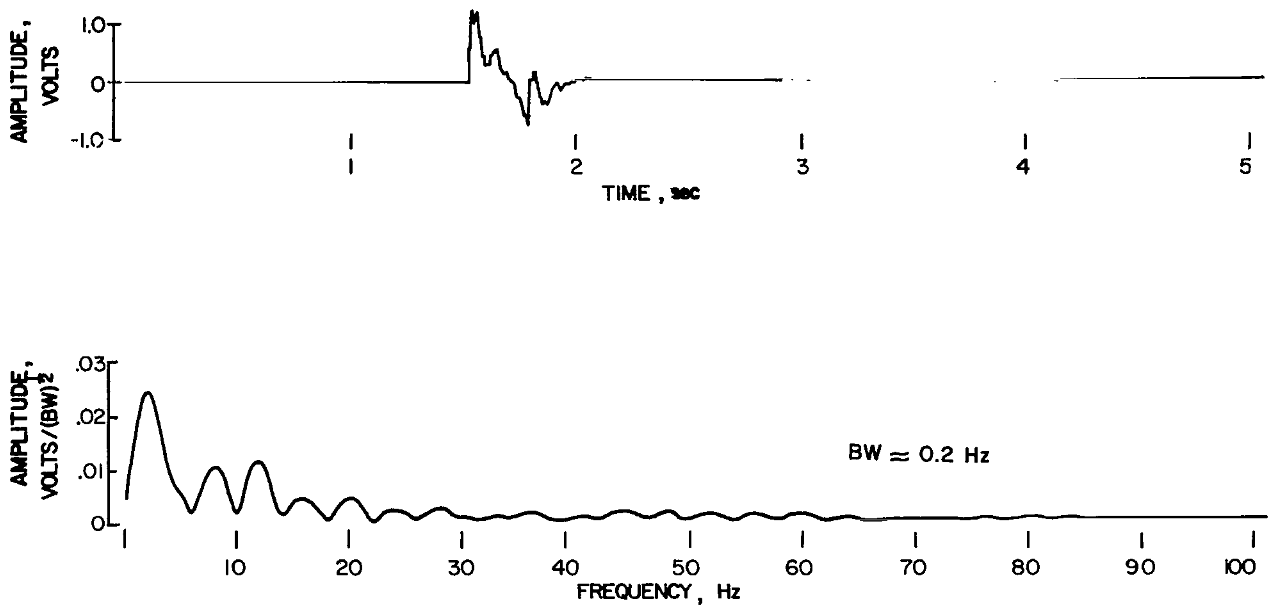

Figure 3.- Amplitude time history and narrow-band spectrum of a sonic boom from a supersonic vehicle.

Implied in the foregoing discussion is the need for an acoustic data acquisition system having response characteristics extending from infrasonic to ultrasonic frequencies. Since the expected amplitude of an acoustic event is not known precisely, an amplitude dynamic range in excess of 70 dB (a factor of 3000) is desirable. In addition to the frequency response and amplitude dynamic range considerations, several operational difficulties arise when making field measurements. First, measurements must be made where the noise sources are located. This necessitates a mobile system which will be subjected to local conditions such as radio and radar emissions, shipboard power, and terrain. Rights-of-way for microphone cables must also be considered. Second, large area coverage is often required to study the "footprint," or area impacted by the noise source; therefore, a large number of simultaneous measurements from dispersed measuring points - and consequently long cable runs and multiple recording stations - are required. Additionally, the local climatic conditions of temperature and humidity affect the measurement systems.

The different measurement systems required to span the total range of interest will, in general, be sensitive - in different ways and to different extents - to the measurement or operational constraints alluded to in the preceding paragraph. Some systems are very sensitive to cable length and type, whereas other systems are sensitive to electromagnetic interference over long cables of any type. Slowly varying atmospheric pressures are not a factor when low frequencies are not of interest, but compensation is required when infrasonic, sonic-boom measurements are made. Some systems, but not all, require additional signal conditioning and ac power or battery packs located close to the microphone station. Generally, present measurement systems require special line drivers or line terminations to send intelligence over long cables.

It is apparent from the preceding discussion that a single measurement system, optimized through modular plug-ins or adapters and minimally affected by the several operational and environmental constraints, is very desirable. This report describes such a system which has recently been developed. The system consists of a condenser microphone, a signal converter, a remotely located signal-conditioning unit (which contains all power, amplification, filtering, control, and calibration circuitry), and peripheral equipment needed to support an acoustic measurement. Experimental results obtained from five production prototype (as opposed to optimized laboratory bench prototype) converter/signal-conditioning units are included along with an operational description of the complete system. A theoretical description of the automatic tuning feature is included in the appendix.

Certain commercial equipment and materials are identified in this paper in order to specify adequately the experimental procedures. In no case does such identification imply recommendation or endorsement of the product by NASA, nor does it imply that the equipment or materials are necessarily the only ones or the best ones available for the purpose. In many cases equivalent equipment and materials are available and would probably produce equivalent results.

SYMBOLS

The following symbols denote mathematical quantities: (a) lowercase letters represent time-varying quantities; (b) uppercase letters with an overbar represent their Laplace transforms; and (c) uppercase letters without an overbar represent reference and quiescent quantities. Values are given in both SI and U.S. Customary Units. The measurements and calculations were made in U.S. Customary Units. Symbols denoting system components used in the figures are not included in this list.

c_m instantaneous microphone capacitance, pF

C_{mQ} quiescent microphone capacitance, pF

C_R varactor diode reference capacitance, pF

C_T total capacitance of 30-dB attenuation network, pF

c_v instantaneous varactor diode capacitance, pF

C_{vQ} quiescent varactor diode capacitance, pF

C_2,C_5 capacitances in differential amplifier-filter, pF

C_6,C_7 capacitances in 30-dB attenuation network, pF

F attenuation factor

$\bar{F}_c$ closed-loop transfer function of automatic tuning control system

$\bar{F}_o$ open-loop transfer function of automatic tuning control system

$\bar{F}_3$ transfer function of differential amplifier-filter

I_R	converter reference current, A
K	$= K_1K_2K_3K_4K_5$
K_1	constant in transfer function of converter, V/pF
K_2	magnitude of Zero Drive® amplifier gain
K_3	constant in transfer function of differential amplifier-filter
K_4	magnitude of gain of summing amplifier
K_5	change in varactor diode network capacitance per volt, pF/V
p	incident sound pressure, Pa
P	constant in transfer function of converter
Q	quality factor of converter tank circuit
R_L	load resistance of converter FET, Ω
R_1 to R_4, R_9 to R_{11}	resistances of feedback branch, Ω
s	complex frequency, used in Laplace transforms, sec^{-1}
v_C	calibration voltage, V
v_O	output voltage of Zero Drive® amplifier, V
V_{R3}	closed-loop reference voltage, V
V_{R4}	open-loop reference voltage, V
v_2	output voltage of converter, V
V_{2Q}	quiescent value of v_2, V
v_3	voltage at test point 3, V
V_{3M}	constant voltage in transfer function of Zero Drive® amplifier, V
V_{3Q}	quiescent value of v_3, V
v_4	output voltage of differential amplifier-filter, V
v_5	varactor diode voltage, V
V_{5Q}	quiescent value of v_5, V
γ	exponent appearing in constant K_5 of varactor diode network

ζ damping ratio of automatic tuning control system

τ_1, τ_2, τ_3 constants in transfer function of differential amplifier-filter, sec

ϕ contact potential (used in expression for constant K_5 of varactor diode network), V

ω_n resonant angular frequency of undamped automatic tuning control system, rad/sec

Abbreviations:

BW bandwidth

CAL calibration

FET field-effect transistor

SPL sound pressure level

DESCRIPTION OF THE UNIFIED SYSTEM

General

The unified acoustic data acquisition system is an AM carrier system consisting primarily of a converter, signal-conditioning electronics, adapters for microphones of various sizes and sound pressure level ranges, and peripheral equipment. Target specifications for the system are as follows:

Microphone, condenser	23.77 mm (1 in.), 12.70 mm (1/2 in.), 6.35 mm (1/4 in.)
Frequency response	0.02 Hz to 20 kHz at the -3-dB points
Frequency bandpass	Selectable by plug-in filters
Amplitude measurement range	50 to 160 dB (ref. 20 μPa)
Dynamic range (single gain setting)	70 dB or greater
Temperature effect upon gain	<2 dB over temperature interval 4° to 54° C (40° to 130° F)
Cable	Two-conductor shielded, twisted pair, up to 600 m (2000 ft) in length
Gain settings	70-dB range with switch increments of 10 dB and 2 dB
Microphone resonant circuit tuning	Automatic or manual
Calibration	By means of remote voltage insertion
Miscellaneous	Indicator for tuning and signal amplitude Output voltage for gain logging

Figure 4 illustrates the unified measurement system concept in its present form.

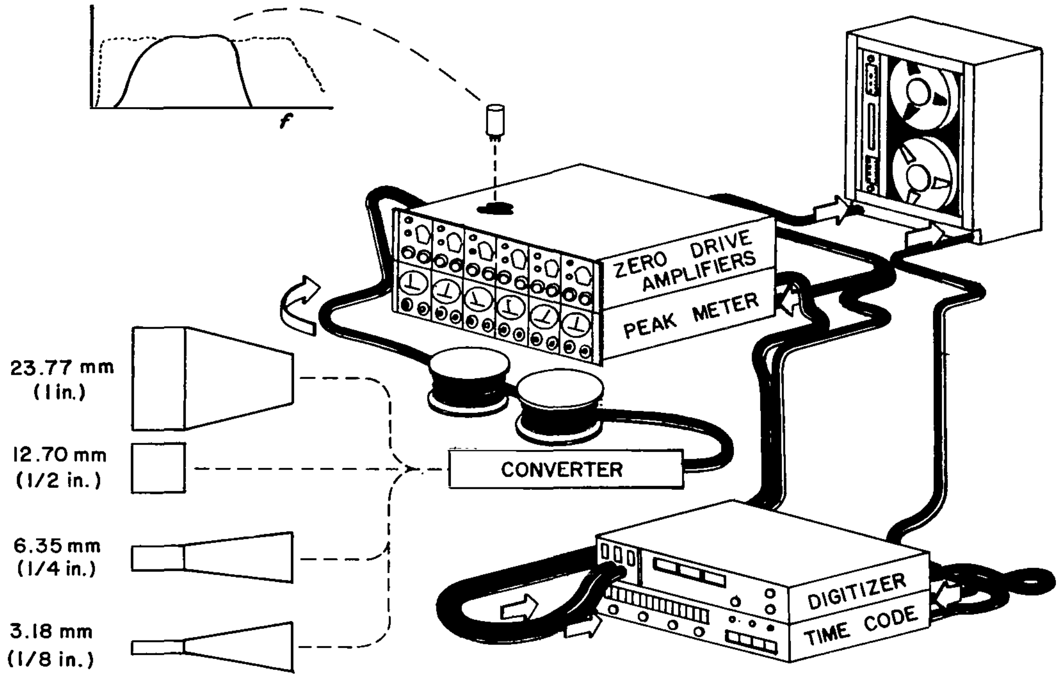

Figure 4.- A conceptual unified acoustic data acquisition system.

The Converter

The function of the converter is to produce an electrical current proportional to the instantaneous sound pressure level at the microphone. Refer to the circuit diagram in figure 5. Transistor T_1 and quartz crystal X make

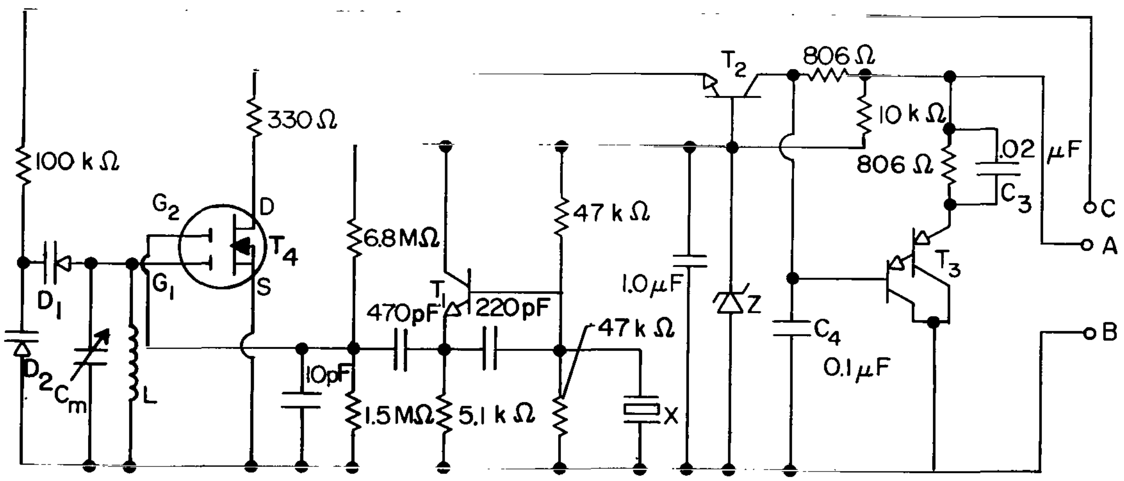

Figure 5.- Circuit diagram of the converter. T_1, 2N3711; T_2, 2N930; T_3, S9120 (Darlington); D_1 and D_2, 1N5463; T_4, 3N202; Z, 1N759A (zener); X, quartz crystal (10 MHz); L, 9-μH inductor. Terminals: A, signal; B, common; C, control.

up the core of the local oscillator, which drives gate G_2 of the field-effect transistor T_4 at 10 MHz. A tank circuit consisting of the condenser microphone C_m, inductor L, and varactor diodes D_1 and D_2 is connected to gate G_1 and tuned to the 10-MHz carrier frequency by means of a direct voltage applied to the control terminal C. As a result of drain-to-gate capacitive coupling, a small fraction of the drain current passes into the tank circuit and produces a voltage at gate G_1. Because the transconductance with respect to gate G_1 (that is, the ratio of drain current to voltage at gate G_1) is exceedingly sensitive to the voltage applied to gate G_2, the signal at gate G_1 mixes with that at gate G_2 to generate a component of direct drain current (over and above the quiescent current). A change in the capacitance C_m changes the level of this direct current; consequently, a periodic change in C_m, as caused by the presence of sound at the microphone, produces a periodic drain current at the frequency of the sound. The drain-to-gate capacitive coupling is usually an unwanted effect in a conventional mixer circuit, but here the effect is used to advantage and permits the detection of very small changes in the capacitance C_m. The output stage, which includes transistor T_2, capacitor C_4, and Darlington transistor T_3, filters out the carrier component of the current and provides a low output impedance for better matching to the low input impedance of the succeeding Zero Drive® amplifier. Capacitor C_3 extends the bandwidth of the converter by increasing the gain of the output stage preferentially for higher acoustical frequencies. A quantitative theory of operation of the converter is given in reference 1.

The circuit elements of the converter are housed in a stainless steel tube 10 cm (4 in.) long × 1.25 cm (1/2 in.) in diameter. One end contains a commercially available connector for the 12.70-mm (1/2-in.) microphone cartridge. Adapters for microphones of other sizes are available. The other end contains a three-terminal connector for the output connections: signal A, common B, and control C.

The Zero Drive® Amplifier[1]

The function of the Zero Drive® amplifier is twofold: (1) to provide power at a constant 22 V dc to the converter terminals A and B, and (2) to amplify the converter signal for recording on an output device. The advantages of the Zero Drive® system include insensitivity to length and type of cable and low triboelectric and other types of noise (ref. 2). Normally an intermediate impedance converter, a line driver, is connected between the signal source and the Zero Drive® amplifier, but here the function of the line driver is transferred to the output stage of the converter. By eliminating the line driver, one is able to use the constant voltage on the line as the supply voltage for the converter without undermining the performance of the Zero Drive® amplifier.

The sequence of stages making up the Zero Drive® amplifier is illustrated in the block diagram of figure 6. The current-to-voltage translator yields an output voltage v_3 which is sensitive to the quiescent as well as the time-varying components of the converter line current (entering terminal A in

[1]Zero Drive® is a new zero impedance signal-conditioning concept for piezoelectric transducers and is a registered trademark of Gilmore Industries, Inc.

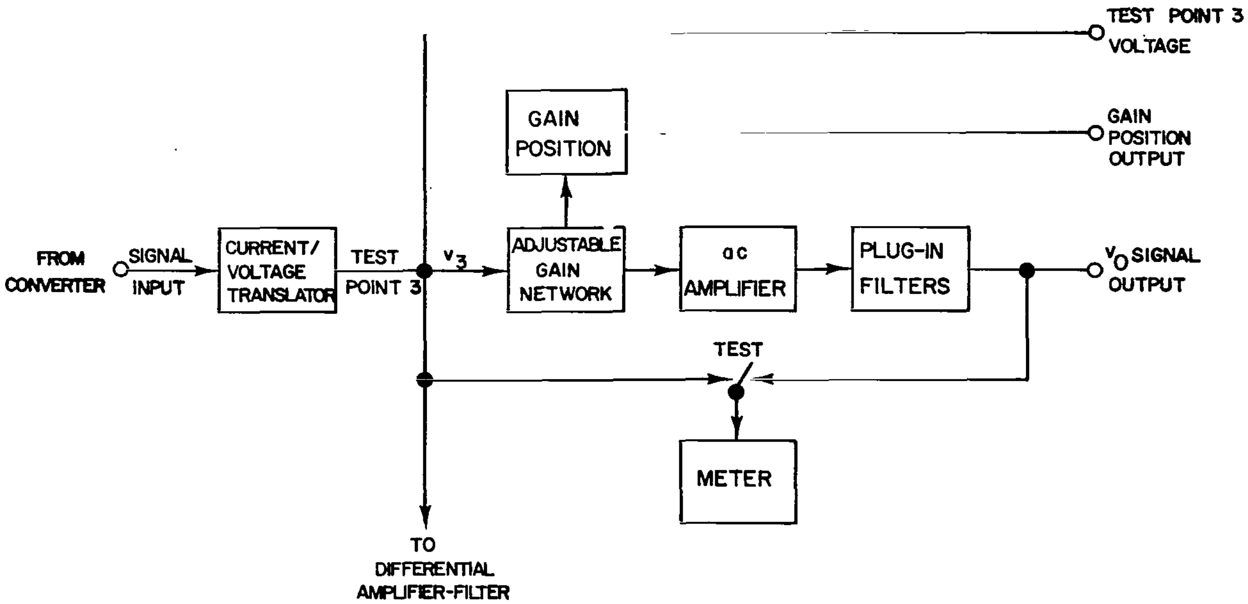

Figure 6.- Block diagram of the Zero Drive® amplifier.

fig. 5). The time-varying component proceeds through the adjustable gain network, ac amplifier, and plug-in filters to the signal output connector. The adjustable gain network contains a double-ganged switch, permitting selection of gain from 0 to 70 dB in 2-dB (fine) and 10-dB (coarse) steps. The bandwidth of the ac amplifier extends from 0.035 Hz to 100 kHz; that of the output signal may be varied from 2, 5, or 10 Hz to 5, 10, or 20 kHz by means of the plug-in filters, where either filter may be bypassed. The meter normally indicates the rms output voltage, but in the test mode it indicates the static component of v_3. The gain position network provides a direct voltage proportional to the settings of the gain position switches.

The output terminal of the current-to-voltage translator, test point 3, is accessible through an external connector. If a low-frequency response down to 0 Hz is desired, as may be the case in sonic-boom measurements, then the quiescent voltage V_{3Q} at test point 3 can be balanced by an external voltage source, and the voltage difference will indicate both static and time-varying changes. Additionally, the quiescent value of v_3, being sensitive to deviations of the converter from optimal tuning, is used as the input voltage in the automatic tuning control system. Figure 7 is a photograph of a five-channel production prototype. Five Zero Drive® amplifiers are situated in a rack together with the power supply.

The Control System

The converter/Zero Drive® system output is excessively sensitive to changes in ambient temperature, primarily because of the pronounced temperature sensitivity of the static microphone capacitance. The net effect of such changes is to detune the converter and reduce the system gain. The automatic tuning control system compensates for changes in static microphone capacitance due to temperature changes and other sources of long-term drift.

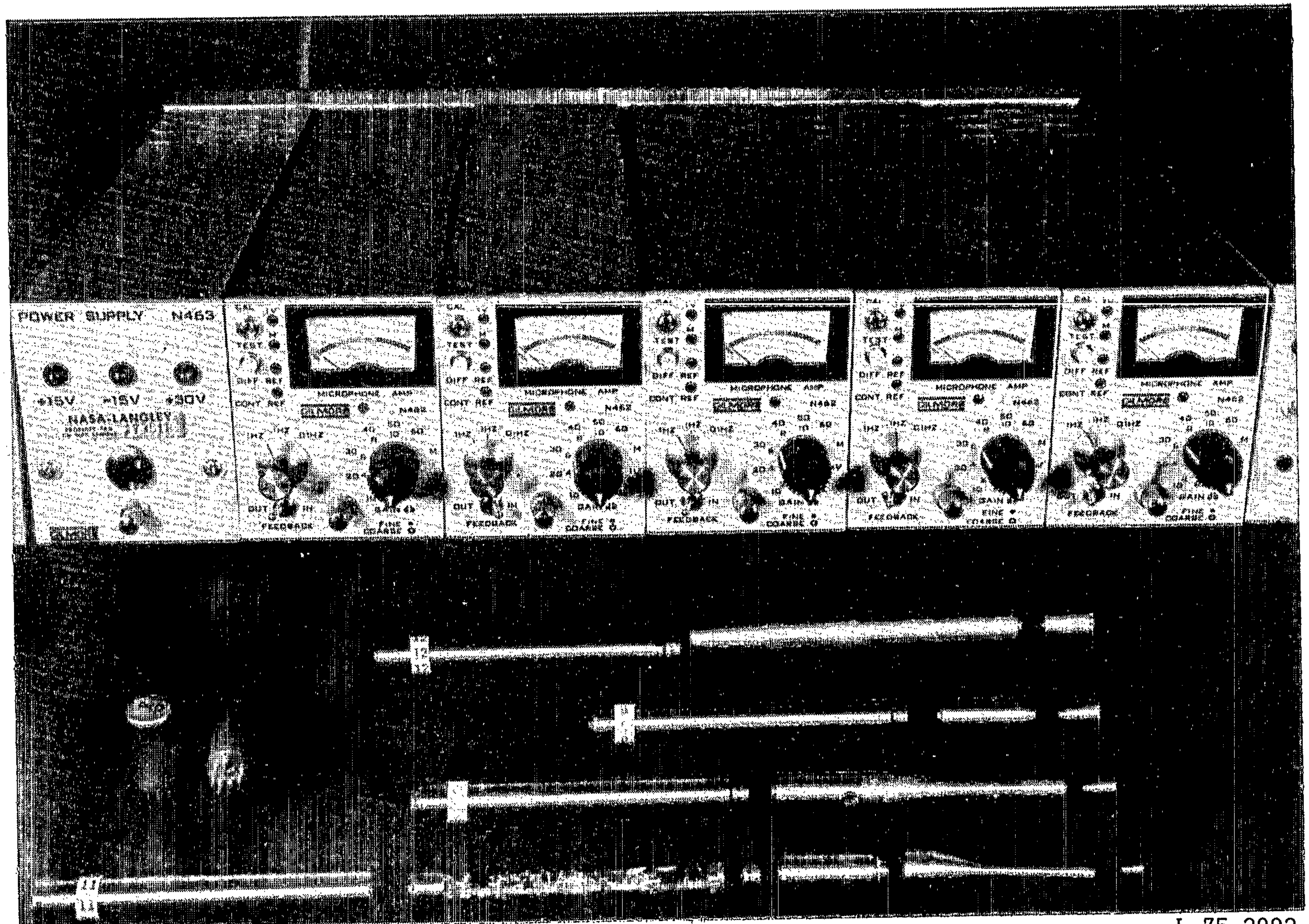

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Figure 7.- Production prototype, including plug-in filters and adapters.

Figure 8 is a block diagram of the converter/Zero Drive® system with automatic tuning control. In the open-loop mode of operation, the differential

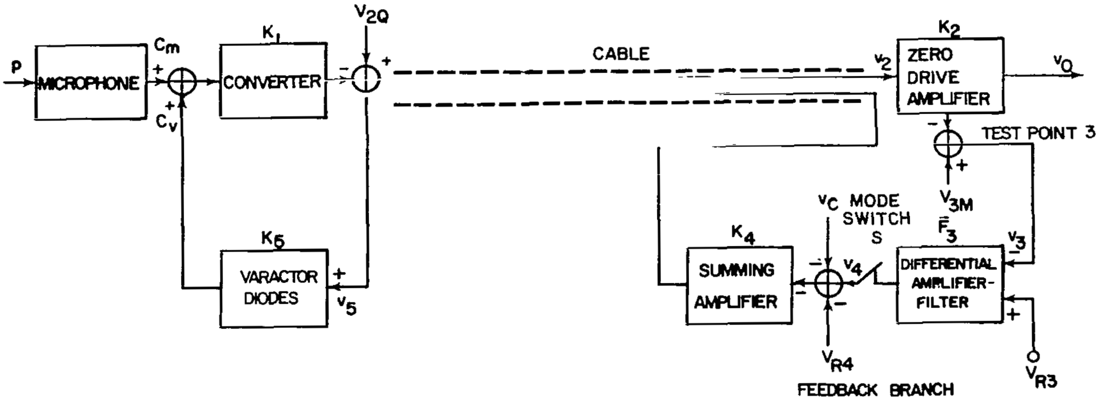

Figure 8.- Block diagram of the converter/Zero Drive® system with automatic tuning control.

amplifier-filter is disconnected from the system by means of switch S. An acoustical sound pressure p produces variations in the microphone capacitance c_m, converter output voltage v_2, and output voltage of the Zero Drive® amplifier v_0, respectively. Manual tuning of the converter is accomplished through adjustment of open-loop reference voltage V_{R4}, which controls the output voltage v_5 of the summing amplifier and thus the capacitance c_v of the varactors.

The system is switched to the closed-loop mode of operation to maintain automatically a fixed converter tuning point. Because the voltage v_3 at test point 3 is highly sensitive to deviations from optimal tuning, the difference between this voltage and the closed-loop reference voltage V_{R3} provides the error voltage which is applied to the feedback branch.

A circuit diagram of the feedback branch, which is built into the existing circuitry of the Zero Drive® amplifier, is shown in figure 9. Under optimal

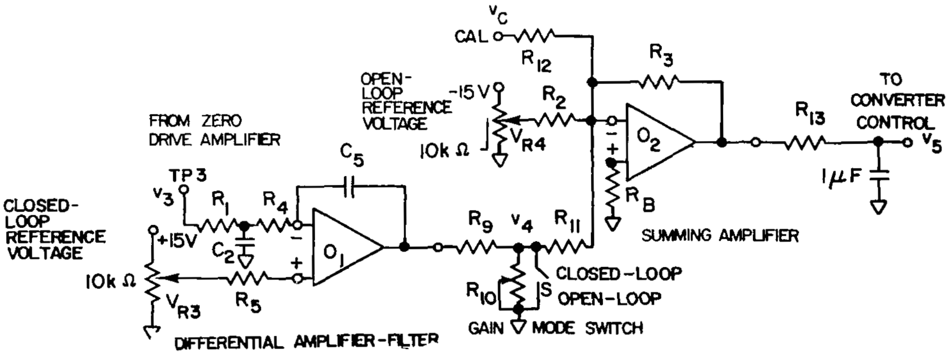

Figure 9.- Circuit diagram of the feedback branch. $R_2 = R_3 = R_{11} = R_{12} = R_{13} = 10$ kΩ; $R_5 = 2$ MΩ; $R_9 = 1$ kΩ; $R_B = 3.3$ kΩ; $R_1 = R_4 = 10$ MΩ at 0.01 Hz, 1 MΩ at 0.1 Hz, 100 kΩ at 1 Hz; $C_2 = 2$ μF; $C_5 = 3.3$ μF; O_1, AD503JH operational amplifier; O_2, AD201AH summing amplifier.

tuning the closed-loop reference voltage V_{R3}, at the noninverting input of operational amplifier O_1, matches the voltage v_3 at test point 3 (TP3), which is applied to the inverting input. The output voltage v_4 of amplifier O_1 is applied to the summing amplifier O_2 through a voltage divider network, which serves as a gain control. The output voltage v_5 of summing amplifier O_2, which determines the capacitance c_v of the varactors, is proportional to the sum of v_4 and the open-loop reference voltage V_{R4}. A change in microphone capacitance c_m, as might be caused by a variation in ambient temperature, produces corresponding changes in voltages v_2, v_3, and v_4. Changes in v_4 produce changes in v_5 and c_v in such a manner as to compensate for changes in c_m. The open-loop reference voltage V_{R4}, however, will not influence the stationary value of v_5, since the controlling action of the system will adjust v_4 to maintain v_5 at the same tuning point at all times, as long as the system remains in the closed-loop mode of operation.

In the open-loop position, mode switch S grounds v_4 and thus prevents the control signal from reaching the summing amplifier O_2. The 1-μF capacitor at the output of O_2 suppresses electrical noise and 60-Hz pickup.

When sound is applied to the microphone, the corresponding variation in v_3 would also normally produce a differential amplifier output, which would cause changes in c_v to compensate for changes in c_m. To prevent cancellation of the acoustical signal, low-pass filtering, provided by resistors R_1 and R_4 and capacitors C_2 and C_5, is built into the differential amplifier stage. The low-pass filter blocks the relatively high acoustical frequencies, but passes the slow variations associated with changes in static capacitance of the microphone in order to permit completion of the control loop. The upper cutoff frequency of the filter can be selected at 0.01, 0.1, or 1 Hz, depending upon the application. The first is used for measurements of sonic-boom signatures, where components down to extremely low frequencies must be rejected; the last for aircraft noise, where the lowest frequency of interest lies around 20 Hz. The selection of cutoff frequency is realized by means of a double-ganged switch on the front panel (not shown in fig. 9), by which the values of both R_1 and R_4 can be simultaneously switched to 10 MΩ, 1 MΩ, or 0.1 MΩ for operation at 0.01 Hz, 0.1 Hz, or 1 Hz, respectively.

In order to calibrate the system, an electrical signal v_c is applied at the input of the summing amplifier. The variations in varactor capacitance caused by the calibration signal produce variations in output voltage similar to those produced by acoustical excitation of the microphone.

The procedure for achieving optimal tuning of the control system is straightforward. First, the system is switched to the open-loop mode of operation, either an acoustical or electrical calibration signal is applied, and the converter is tuned by adjusting V_{R4} to obtain a maximum output voltage v_0. Then the system is switched to the closed-loop mode and the procedure is repeated with an adjustment of V_{R3}. An alternative to monitoring output voltage v_0 is to press the TEST button on the front panel and to adjust V_{R4} and V_{R3} for a prescribed reading on the panel meter.

A theoretical analysis of the automatic tuning control system is given in the appendix.

Physical Description

The system controls are located on the front panel of the Zero Drive® amplifier. (See fig. 7.) These include coarse and fine gain switches in 10-dB and 2-dB steps, respectively; a test button to select the meter indication as quiescent voltage at test point 3 or as rms output voltage; a calibration jack for introducing calibration voltages into the amplifier and meter circuits; potentiometers to set the closed-loop and open-loop reference voltages; a feedback mode switch to select the closed-loop or open-loop mode of operation; and a switch to select the upper cutoff frequency of the automatic tuning control system to 0.01, 0.1, or 1 Hz.

The power supply provides 30 V, 15 V, and -15 V, within ±1 percent, to all five Zero Drive® amplifiers in the rack. Light-emitting diodes indicate when the supply voltages are 2 percent or more out of tolerance.

Adapters

Adapter for 30-dB attenuation.- A diagram of the attenuation network is shown in figure 10. The attenuation factor F and total tank circuit

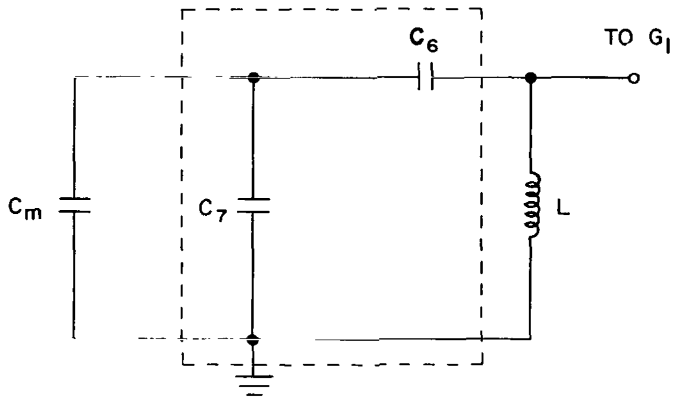

Figure 10.- Network for 30-dB attenuation.

capacitance C_T are given by the following expressions:[2]

$$F = \frac{(C_6 + C_7 + C_m)(C_7 + C_m)}{C_6 C_m}$$

$$C_T = \frac{C_6(C_7 + C_m)}{C_6 + C_7 + C_m}$$

[2]The attenuation factor F relates the relative change in total tank circuit capacitance to the relative change in microphone capacitance alone:

$$\frac{\delta C_T}{C_T} = \frac{1}{F}\frac{\delta C_m}{C_m}$$

Values for C_6 and C_7 are chosen to yield the desired value of F and to insure that $C_T \approx C_m$ for the purpose of maintaining tuning at the given carrier frequency. Using the values $C_6 = 22$ pF, $C_7 = 100$ pF, $C_m = 20.5$ pF (typical value) gives F = 38 (31.6 dB) and $C_T = 18.6$ pF. Thus F lies very close to the desired attenuation factor, and C_T lies within the interval of tuning.

The capacitors C_6 and C_7 are housed in a commercially available 12.70-mm (1/2-in.) dehumidifier (with desiccant removed), which can be screwed directly between the microphone and the converter. To connect C_6 to the network requires a mechanical modification to the dehumidifier whereby the dehumidifier pin has to be severed.

Adapter for 23.77-mm (1-in.) microphone.- Because the capacitance of a 23.77-mm (1-in.) microphone (≈45 pF) exceeds that of 12.70-mm (1/2-in.) microphone (≈20 pF), the inductance of the tank circuit must be reduced to permit the converter to be tuned at the same carrier frequency in both cases. For this purpose, a parallel 9-μH inductor is introduced into the tank circuit. The parallel inductor is housed in a similar 23.77-mm (1-in.) dehumidifier (with desiccant removed), which is screwed between the microphone and a mechanical adapter; the latter can be screwed directly onto the converter. Automatic tuning control is readily achieved, as in the case of the 12.70-mm (1/2-in.) microphone.

Adapter for 6.35-mm (1/4-in.) microphone.- The 6.35-mm (1/4-in.) microphone, on the other hand, has a smaller capacitance (≈6.5 pF) than the 12.70-mm (1/2-in.) microphone. In order to maintain the converter tuning point, an inductor (with an inductance of ≈15 μH) is placed in series with the condenser microphone. The inductor is located in a special housing between the 6.35-mm (1/4-in.) adapter and the microphone.

Peripheral Equipment

The unified acoustic data acquisition system, as illustrated in figure 4, includes such additional support equipment as peak signal indicators, a multiplexer, an analog-to-digital converter, a time code generator, and an FM tape recorder.

Tape recorders are used to store the acoustic data acquired at remote measurement locations for later processing in a central computer facility. The tape recorders used in this system are of the frequency modulation type required to provide a frequency response from 0 Hz to 20 kHz and a broadband dynamic amplitude range in excess of 50 dB.

If an FM tape channel is overdriven, total loss of data results. Hence, a means for detecting that an overdriven situation has occurred, and when it occurred, is highly desirable for use during the data analysis phase. Also, system gain settings and other housekeeping information, suitably recorded, are valuable during the analysis phase. Provision for doing these tasks has been made in this system through the time-division multiplexing of voltages proportional to the gain switch settings and to the outputs of peak signal detection circuits for subsequent analog-to-digital conversion. The 45-control-bit portion of the time code frame is then used to merge these data with the standard time code information. Once this information and the acoustic analog signals are recorded on tape, the concept of a single, unified system for all measurement tasks is complete, with all needed information conveniently and efficiently accounted for. Suitable interfacing with computer facilities completes an efficient system for acquiring and processing acoustic data.

SYSTEM PERFORMANCE

All laboratory tests described in this section were performed with the following system components: microphone, 12.70-mm (1/2-in.), condenser (pressure type), with dessicator; amplifier, Zero Drive®, with high-pass bypass and low-pass 20-kHz filters; converter, number 8 (laboratory prototype); and cable, two 0.76-mm diameter (22 AWG) stranded conductors, twisted pair, aluminum foil shielded. The exceptions include the use of converter number 5 for the temperature tests and the 23.77-mm (1-in.) and 6.35-mm (1/4-in.) microphone cartridges in conjunction with their respective adapters.

Dynamic Range

The useful operating range of amplitude in an acoustical data acquisition system is limited by noise at low levels and by signal distortion at high levels. The dynamic range of a system is defined here to extend from 5 dB above the electrical noise floor to the sound pressure level at 4 percent distortion for a pure tone.

The following procedure was used to determine the noise floor of the system:

(1) An acoustic calibrator was used to apply a 100-dB sound pressure level (SPL) to the microphone at 1 kHz. The converter was optimally tuned, and the output voltage of the Zero Drive® amplifier was measured on a 1/3-octave analyzer.

(2) A calibration voltage v_C at 1 kHz inserted at the calibration jack ("CAL," fig. 9) was adjusted to produce the same output voltage as in step 1 and was thus made equivalent to the 100-dB acoustical signal.

(3) The converter microphone was replaced by a dummy microphone having approximately the same capacitance. The calibration voltage, as adjusted under step 2, was reapplied, and the converter retuned to produce an output voltage corresponding to the 100-dB SPL on the 1/3-octave analyzer. This adjustment made the gain of the system with the dummy microphone equal to that of the system with the actual microphone.

(4) Upon removal of the calibration voltage, the wide-band output noise (22.4 Hz to 22.4 kHz) was measured on the 1/3-octave analyzer.

Because the noise floor was found to be dependent upon the range setting of the Zero Drive® amplifier, the results for two range settings[3] are shown in table I. Except for the 60-Hz component and its harmonics, the noise is white noise and increases linearly with 1/3-octave-band number.

TABLE I.- DYNAMIC RANGE OF CONVERTER/ZERO DRIVE® SYSTEM

FOR VARIOUS CABLE LENGTHS

Cable length, m (ft)	Range setting of Zero Drive® amplifier	Noise floor, dB (a, b)	SPL at 4 percent distortion, dB (c)	Dynamic range, dB (d)
1	20	56	137	76
(3)	30	50	127	72
300	20	56	135	74
(1000)	30	51	126	70
600	20	56	133	72
(2000)	30	50	125	70
900	20	56	132	71
(3000)	30	50	125	70

[a]Bandwidth of noise floor measurement: 22.4 Hz to 22.4 kHz.
[b]Reference SPL: 20 μPa.
[c]Frequency of distortion measurement: 1 kHz.
[d]5 dB above noise floor to SPL at 4 percent distortion.

The distortion of the output voltage of the Zero Drive® amplifier was measured on a distortion analyzer. For this measurement the system was excited by application of a calibration voltage because of an inherently high distortion in the acoustic calibrator at high sound pressure levels (>140 dB). For all measurements, the frequency was 1 kHz. In table I, various range settings and cable lengths are listed for the noise floor, for the highest sound pressure level at which the distortion does not exceed 4 percent, and for the corresponding dynamic range.

As is evident from table I, the SPL at 4 percent distortion decreases slightly with increasing cable length. This effect is associated with the slight increase in system gain with cable length, about 0.8 dB per 300 m (1000 ft) of cable. However, this added gain does not decrease the noise floor, because of noise generated in the additional length of cable. From the relative constancy of the noise floor and dynamic range, it is concluded that the Zero Drive® system is capable of driving very long lengths of cable.

[3]On other range settings, the dynamic range is somewhat less than 70 dB, the minimum needed to meet specifications.

Frequency Response

An electrostatic actuator, driven at an equivalent SPL of 90 dB, was used to determine the frequency response of the system. The frequency dependence of the output of the Zero Drive® amplifier is shown in figure 11 for four cable

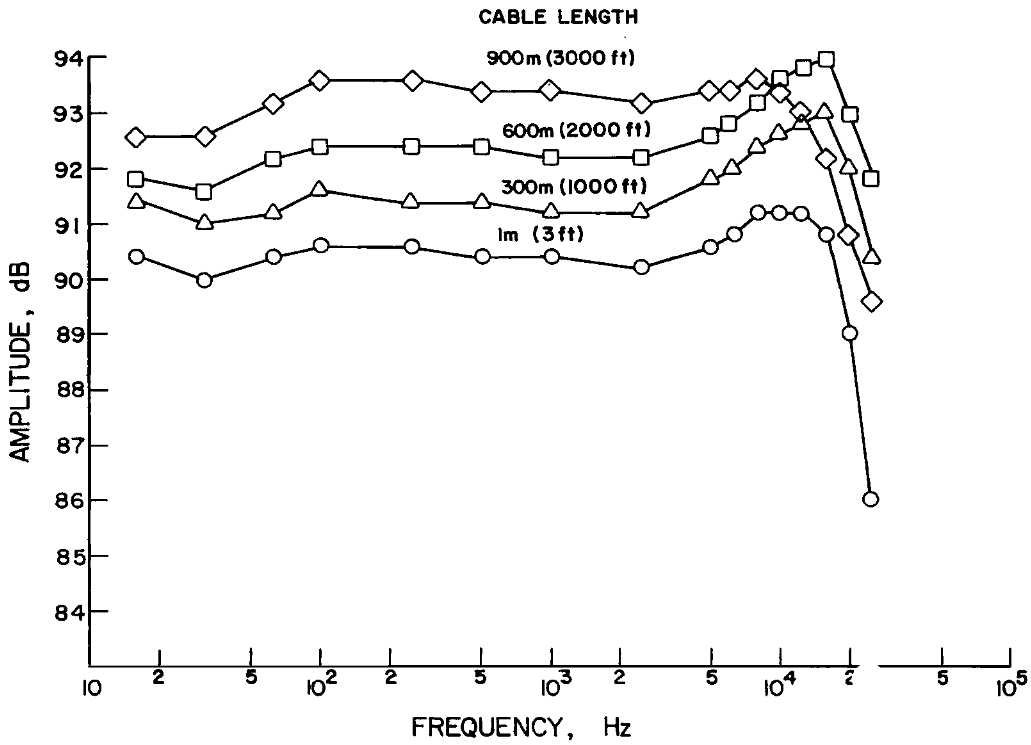

Figure 11.- Frequency response of converter/Zero Drive® system with 12.70-mm (1/2-in.) pressure microphone and with Zero Drive® amplifier set on range 30. Filters: low-pass bypass and high-pass bypass.

lengths up to 900 m (3000 ft). In order to meet the specification for cable lengths up to 900 m (3000 ft), it was necessary to reduce the input resistance of the Zero Drive® amplifier from 50 Ω to 10 Ω.

The capacitor C_3 in the converter circuit of figure 5 serves to compensate for premature rolloff at higher audio frequencies. The value of C_3 sufficiently compensates at 900 m (3000 ft) of cable, overcompensates slightly at 1 m (3 ft), and overcompensates by as much as 2 dB at 300 m (1000 ft) and 600 m (2000 ft). In all instances the amplitude remains within a 3-dB band over the entire operating frequency range.

Adapters

The system performance with the 30-dB attenuator and the 23.77-mm (1-in.), and 6.35-mm (1/4-in.) adapters is summarized in tables II, III, and IV, respectively. Automatic tuning control is readily achieved with all adapters.

TABLE II.- RESPONSE OF THE CONVERTER/ZERO DRIVE• SYSTEM

WITH 30-dB ATTENUATOR

[Attenuator: C_6 = 22 pF; C_7 = 100 pF (see fig. 10)
Cable length: 1 m (3 ft)]

Range setting of Zero Drive• amplifier	Noise floor, dB (a,b)	SPL at 4 percent distortion, dB (c)	Dynamic range, dB (d)	Attenuation, dB
20	90	160	65	32.6
30	82	159	72	32.6

[a]Bandwidth of noise floor measurement: 22.4 Hz to 22.4 kHz.
[b]Reference SPL: 20 μPa.
[c]Frequency of distortion measurement: 1 kHz.
[d]5 dB above noise floor to SPL at 4 percent distortion.

TABLE III.- RESPONSE OF THE CONVERTER/ZERO DRIVE• SYSTEM

WITH 23.77-mm (1-in.) ADAPTER

[Microphone: 23.77-mm (1-in.) condenser microphone (very low frequency pressure type)
Parallel inductance in adapter: 9 μH
Cable length: 1 m (3 ft)]

Range setting of Zero Drive• amplifier	Noise floor, dB (a,b)	SPL at 4 percent distortion, dB (c)	Dynamic range, dB (d)
10	55	122	62
20	44	122	73
30	41	113	67

[a,b,c,d]See footnote to table II.

TABLE IV.- RESPONSE OF THE CONVERTER/ZERO DRIVE• SYSTEM

WITH 6.35-mm (1/4-in.) ADAPTER

[Microphone: 6.35-mm (1/4-in.) condenser microphone and adapter
Series inductance in adapter: 15 μH
Cable length: 1 m (3 ft)]

Range setting of Zero Drive• amplifier	Noise floor, dB (a,b)	SPL at 4 percent distortion, dB (c)	Dynamic range, dB (d)
20	75	150	70
30	68	142	69

[a,b,c,d]See footnote to table II.

Temperature Effects

The most crucial test of the automatic tuning control is to determine how well the system maintains optimal tuning over the specified temperature interval 4° to 54° C (40° tc 130° F). For this test the converter was placed in a controlled temperature chamber, and the response to a calibration voltage at constant amplitude was measured as a function of temperature. The calibration voltage was adjusted to produce a signal equivalent to an SPL of 100 dB at 1 kHz at the output of the Zero Drive® amplifier. The output voltage, the tuning control voltage, and the voltage at test point 3 were recorded at selected temperatures.

The variation of output amplitude with temperature is shown in figure 12 for both the open-loop and closed-loop modes. The converter was tuned for

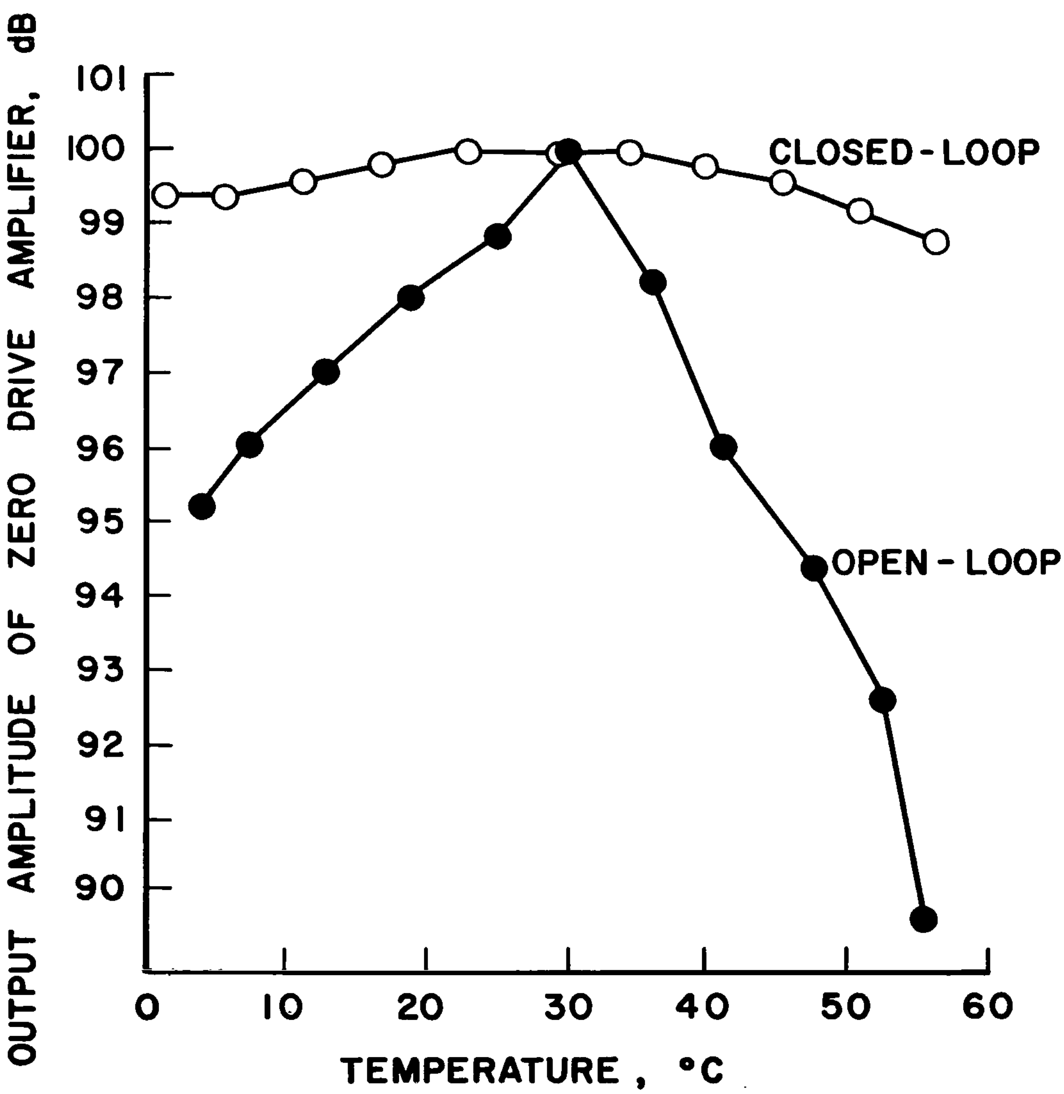

Figure 12.- Temperature sensitivity of converter/Zero Drive® system. Converter input level, 100 dB; range setting of Zero Drive® amplifier, 20.

maximum output voltage at the reference temperature of 30° C (86° F). The drop in open-loop sensitivity on either side of the maximum is attributable to the detuning associated with the change in microphone capacitance with temperature. The figure shows that the automatic tuning control system successfully maintains the system gain to within ±1 dB over the entire temperature interval.

As shown in figure 13, the quiescent varactor voltage V_{5Q} increases with temperature. This increase causes varactor capacitance c_v to decrease and

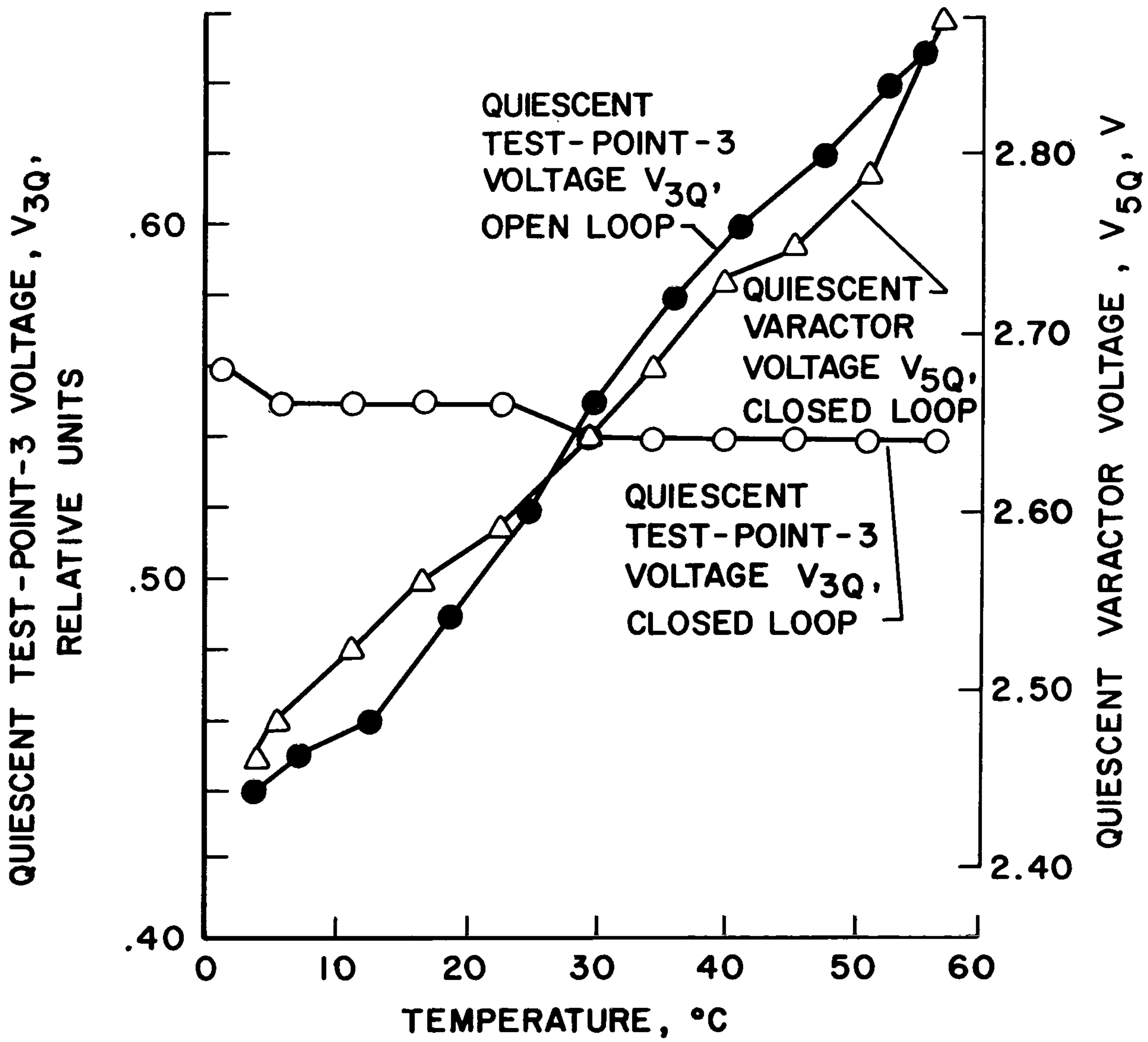

Figure 13.- Temperature dependence of quiescent varactor voltage V_{5Q} and quiescent voltage V_{3Q} at test point 3.

thereby compensate for the increase in microphone capacitance. In the open-loop mode, the quiescent voltage V_{3Q} at test point 3 follows changes in microphone capacitance; in the closed-loop mode, however, the automatic tuning control is effective in keeping V_{3Q} closely clamped to the reference voltage V_{R3}.

FIELD TEST

The unified acoustic data acquisition system as described in this report, configured with a 12.70-mm (1/2-in.) microphone, was deployed in late November 1975 for a helicopter-flyover field exercise. Data for several level flyovers at an altitude of 75 m (250 ft) were obtained by using two converter systems on 300-m (1000-ft) cables. Figure 14 is an acoustic data time history and the

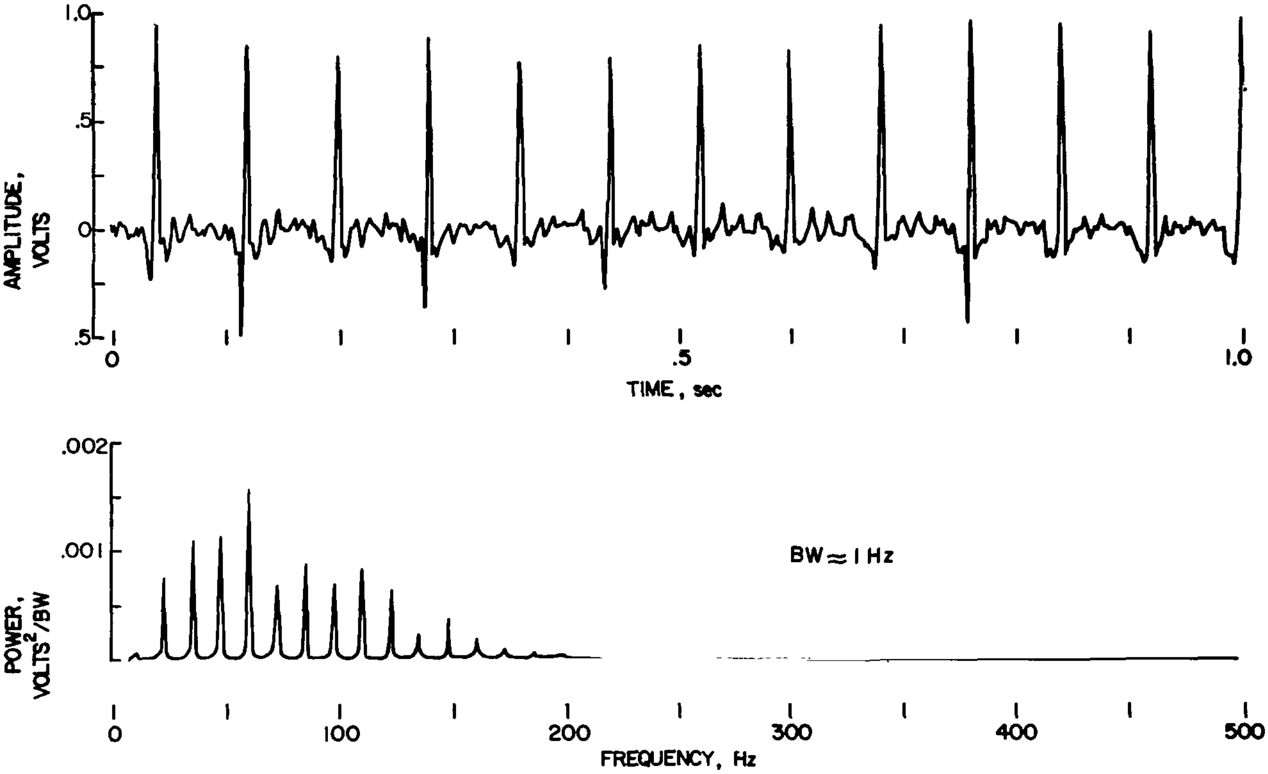

Figure 14.- Amplitude time history and narrow-band power spectrum of a turbine-powered helicopter obtained with the developed system.

associated power spectrum from one flyover. The power spectrum shows evidence of significant blade slap and, like the time history, is seen to be similar in dynamic characteristics to the data of figure 2 obtained with a previous system.

CONCLUDING REMARKS

A unique, unified system for the acquisition of acoustic data has been designed, developed, and tested. The system is readily adaptable to the common condenser microphone sizes; has a frequency response extending from 0 Hz to 20 kHz at the -3-dB point; has selectable frequency bandpass through plug-in filter modules; has a dynamic range of 70 dB; has a variation in sensitivity due to temperature of less than ± 1 dB over the temperature range of 4° to 54° C (40° to 130° F); is relatively insensitive to cable length, having been tested on a shielded, two-conductor, twisted-pair cable up to 900 m (3000 ft) in length; has a 70-dB voltage gain range selectable in 2-dB increments; has an open-loop (manual) and a closed-loop (automatic) remote tuning capability; features a remote electrical calibration capability; and contains a meter to monitor the signal level or the quiescent tuning point. These features satisfy target specifications.

Langley Research Center
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February 3, 1977

APPENDIX

THEORETICAL ANALYSIS OF THE AUTOMATIC TUNING CONTROL SYSTEM

Transfer Functions of System Components

Microphone-converter.- According to equations (10), (13), and (18) of reference 1, the output voltage v_2 of the converter (see fig. 8) is related to the capacitance of the tank circuit as follows:

$$v_2 = V_{2Q} - K_1(\delta c_m + \delta c_v)$$

where[4]

V_{2Q} quiescent value of v_2

δc_m $\equiv c_m - C_{mQ}$

δc_v $= c_v - C_{vQ}$

c_m instantaneous microphone capacitance

c_v instantaneous varactor capacitance

C_{mQ} quiescent microphone capacitance

C_{vQ} quiescent varactor capacitance

The constant K_1 is made up of the following converter parameters:

$$K_1 = \frac{I_R R_L P Q^2}{C_{mQ} + C_{vQ}}$$

in which

I_R converter reference current (see eq. (12), ref. 1)

R_L load resistance of converter FET (330-Ω resistor of fig. 5)

P ≈ 1 (see eq. (11), ref. 1)

Q quality factor of tank circuit

In the complex frequency s domain,

$$\bar{V}_2 = \frac{V_{2Q}}{s} - K_1(\delta\bar{C}_m + \delta\bar{C}_v) \tag{1}$$

---

[4]See Symbols for further explanation of convention used.

APPENDIX

<u>Zero Drive® amplifier</u>.- For the purpose of analysis, the transfer function of the Zero Drive® amplifier includes the output stage of the converter, which has a gain somewhat greater than unity. The relationship between voltage v_3 at test point 3 and output voltage v_2 may be described by the following expression over the linear range:

$$v_3 = V_{3M} - K_2 v_2$$

and in the s domain

$$\bar{V}_3 = \frac{V_{3M}}{s} - K_2 \bar{V}_2$$

Equation (1) is used to relate $\bar{V}_3$ to the capacitance changes:

$$\bar{V}_3 = \frac{V_{3Q}}{s} + K_1 K_2 \left(\delta\bar{C}_m + \delta\bar{C}_v\right) \tag{2}$$

where $V_{3Q} = V_{3M} - K_2 V_{2Q}$ is the quiescent value of v_3. The values of the constants in equation (2) have been determined experimentally and are

$$V_{3Q} = 9.20 \text{ V}$$

$$K_1 = 4.3 \text{ V/pF}$$

$K_2 \approx 3.6$, with Zero Drive® amplifier on 30-dB range

<u>Differential amplifier-filter</u>.- The output voltage $\bar{V}_4$ is related to $\bar{V}_3$ and V_{R3} in the following manner:

$$\bar{V}_4 = \bar{F}_3\left(\bar{V}_3 - V_{R3}\,\frac{1 + \tau_3 s}{s}\right) \tag{3}$$

In equation (3) appear the transfer function, time constants, and gain:

$$\bar{F}_3 = -\frac{K_3}{\tau_1^2 s^2 + \tau_2 s}$$

$$\tau_1 = \left(R_1 R_4 C_2 C_5\right)^{1/2}$$

$$\tau_2 = \left(R_1 + R_4\right) C_5$$

$$\tau_3 = R_1 C_2$$

$$K_3 = \frac{R_{10}}{R_9 + R_{10}}$$

Later in the analysis the parameters τ_1, τ_2, τ_3, and K_3 will be chosen to achieve the specified cutoff frequency and damping ratio of the closed-loop system.

APPENDIX

Summing amplifier.- The relationship between output and input is simply

$$\bar{V}_5 = -K_4\left(\bar{V}_4 + \frac{V_{R4}}{s}\right) \tag{4}$$

and

$$K_4 = \frac{R_3}{R_2} = \frac{R_3}{R_{11}} = 1$$

Varactor.- The relationship between the capacitance c_v and voltage v_5 of a varactor may be approximated by the following expression:

$$c_v = \frac{C_R}{\left(1 + \frac{v_5}{\phi}\right)^{\gamma}}$$

where

C_R varactor capacitance at zero bias

γ diode power-law exponent (≈ 0.44)

ϕ contact potential (≈ 0.6 V)

This relationship is linearized for small changes about the quiescent point (V_{5Q}, C_{vQ}):

$$c_v = C_{vQ} - K_5(v_5 - V_{5Q}) \tag{5}$$

Taking into account the series arrangement of the two varactors in figure 5, the constant K_5 is determined as follows:

$$K_5 = -\frac{1}{2}\left(\frac{\partial c_v}{\partial v_5}\right)_{V_{5Q}} = \frac{\gamma C_R}{2\phi\left(1 + \frac{V_{5q}}{\phi}\right)^{\gamma+1}}$$

If the values of γ and ϕ given previously and the typical values $C_R = 22$ pF (type 1N5463A) and $V_{5Q} = 2$V are used, then $\frac{1}{2} C_{vQ} = 5.8$ pF (series combination) and $K_5 = 0.977$ pF/V.

Closed-Loop Transfer Function

The system of equations (1) to (5) is sufficient to specify the behavior of the closed-loop control system. In particular, the change in varactor capacitance becomes

$$\delta\bar{C}_v = \frac{K_3K_4K_5\tau_3V_{R3} + (\tau_1{}^2s + \tau_2)K_5(K_4V_{R4} + V_{5Q}) - K\delta\bar{C}_m}{\tau_1{}^2s^2 + \tau_2 s + K} \tag{6}$$

where

$$K = K_1K_2K_3K_4K_5$$

For the derivation of equation (6), the following relationship is used:

$$V_{3Q} = V_{R3} \tag{7}$$

In order to prove equation (7), consider the case in which the converter is optimally tuned:

$$\delta\bar{C}_v + \delta\bar{C}_m = 0 \tag{8}$$

The steady-state value of the output of the differential amplifier-filter is found by applying the final-value theorem of Laplace transform theory to equation (3):

$$v_4(\infty) = \lim_{s\to 0} s\bar{V}_4 = \lim_{s\to 0} K_3\left(-\frac{s\bar{V}_3 - V_{R3}}{\tau_1{}^2s^2 + \tau_2 s} + \frac{V_{R3}\tau_3}{\tau_1{}^2 s + \tau_2}\right)$$

Now, since by equation (2)

$$\lim_{s\to 0} s\bar{V}_3 = v_3(\infty) = V_{3Q}$$

it follows that the first limit above can exist if, and only if, equation (7) is true. The second term yields

$$v_4(\infty) = \frac{K_3\tau_3V_{R3}}{\tau_2} = \frac{K_3R_1C_2V_{R3}}{(R_1 + R_4)C_5}$$

This voltage less the reference voltage V_{R3} is the stationary voltage across capacitor C_5.

The inverse transforms of the V_{R3}, V_{R4}, and V_{5Q} terms in equation (6) yield exponentially decaying time functions, which vanish rapidly after the system power is turned on. The closed-loop transfer function, then, is taken to be the following:

$$\bar{F}_c(s) = \frac{\delta\bar{C}_v}{\delta\bar{C}_m} = -\frac{K}{\tau_1{}^2s^2 + \tau_2 s + K} \tag{9}$$

A Nyquist plot of the open-loop transfer function

$$\bar{F}_o(s) = \frac{K}{\tau_1{}^2s^2 + \tau_2 s}$$

reveals that the system is always stable as long as $K > 0$.

APPENDIX

Theoretical Results

Steady-state error.- The steady-state response of the varactor capacitance to a step change in microphone capacitance $\delta\bar{C}_m/s$ is, from equation (9),

$$\delta c_v(\infty) = \lim_{s\to 0} s\delta\bar{C}_v = \lim_{s\to 0} \frac{-Ks}{\tau_1{}^2 s^2 + \tau_2 s + K} \frac{\delta\bar{C}_m}{s} = -\delta\bar{C}_m$$

In other words, the change in c_v compensates exactly for the change in c_m, and the system responds with zero steady-state error. Applying this procedure to equation (6) instead of equation (9) shows that this result is independent of the reference voltage V_{R4}. Experimentally, it has been found that the system always approaches the optimal tuning point, independently of the setting of V_{R4}, if the tank circuit capacitance is within the linear range of operation of the converter. After the system power is turned on, only a slight adjustment of V_{R3} is generally needed to tune the converter.

Design criteria.- The closed-loop transfer function given by equation (9) can be written in the form of a second-order resonant system:

$$\bar{F}_c(s) = \frac{-\omega_n{}^2}{s^2 + 2\zeta\omega_n s + \omega_n{}^2}$$

where $\omega_n = \sqrt{K}/\tau_1$ is the resonant angular frequency of the undamped system, and $\zeta = \tau_2/2\tau_1\sqrt{K}$ is the damping ratio. The magnitude of $\bar{F}_c$ is unity at zero frequency and is down 6 dB at approximately ω_n if the damping is critical. It is desired that (a) $f_n = \omega_n/2\pi = 0.01$ Hz (minimum upper cutoff frequency), and (b) $\zeta \approx 1$ (critical damping). These values are attained approximately with the values of circuit components R_1, R_4, C_2, and C_5 given in the legend of figure 9 together with $K \approx 3$. The latter is adjusted by means of rheostat R_{10}. In order to accommodate the high-input resistors $R_1 = R_4 = 10$ MΩ, it is necessary that 0_1 be an FET operational amplifier, in this case a type AD503JH, having an input resistance exceeding 100 GΩ. The upper cutoff frequency may be raised to 0.1 or 1 Hz simply by lowering both R_1 and R_4 to 1 MΩ or 0.1 MΩ, respectively.

Calibration signal.- If a calibration voltage v_C is applied to the input of the summing amplifier (see fig. 8), then the output voltage $\bar{V}_0$ of the Zero Drive® amplifier is the following:

$$\bar{V}_0 = \frac{K_1K_2(\tau_1{}^2 s^2 + \tau_2 s)(K_4K_5\bar{V}_C + \delta\bar{C}_m)}{\tau_1{}^2 s^2 + \tau_2 s + K}$$

The calibration voltage is equivalent to a change in microphone capacitance equal to $K_4K_5\bar{V}_C$, in the Laplace transform notation.

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